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SATELLITE BROADCASTING:

possible advantages of using digital modulation for television

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Research Department, Engineering Division
THE BRITISH BROADCASTING CORPORATION

RESEARCH DEPARTMENT

SATELLITE BROADCASTING: POSSIBLE ADVANTAGES OF USING DIGITAL MODULATION FOR TELEVISION

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SATELLITE BROADCASTING: POSSIBLE ADVANTAGES OF USING DIGITAL MODULATION FOR TELEVISION

Summary

This report is concerned with the broadcasting of television from satellites. It compares the carrier-to-noise ratios and co-channel protection ratios required for digitally-coded modulation, with those required for frequency modulation.

It is unlikely that a digital system will permit a significantly lower transmitter power than that required for a f.m. system. However, a 12-phase digital system may have an advantage of about 2 to 8 dB from the point of view of co-channel interference.

1. Introduction

There has recently been a rapid increase in the application of digital coding techniques to communication systems, and it is desirable to consider if such techniques would be advantageous for direct broadcasting of a television signal from a satellite to a domestic receiving installation.

Existing 625-line television receivers should be capable of receiving the transmission with as simple an adaptor as possible. Low cost and simplicity of the circuits in the receiver are of prime importance, and this must strongly influence the choice of system.

In order to study the problem, a number of basic decisions must be taken, concerning such parameters as the available bandwidth, and the acceptable picture quality. These matters are still the subject of national and international discussion, but for the purpose of this report, the following assumptions will be made:

(a) The television baseband signal is of the U.K. (Standard I) 625-line form and occupies a bandwidth of 5·5 MHz. We shall not consider such possibilities as transmitting the luminance and chrominance components as separate signals, or of transmitting the sound information during the synchronising pulse period.

(b) For each vision channel there is only one audio channel of up to 15 kHz bandwidth. There may be advantages in having more audio channels available for multilingual broadcasts, additional sound programmes, control signals etc., but in the absence of specific proposals, these requirements will not be considered further in this report.

(c) The r.f. bandwidth of each vision channel, together with its sound channel, will be taken as between 24 and 32 MHz.[1]

2. Required signal-to-noise ratios at receiver output

For the vision signal, the signal-to-noise ratio (s.n.r.) at the receiver output is defined as

$$20 \log_{10} \left\{ \frac{\text{peak-to-peak amplitude of picture luminance}}{\text{r.m.s. noise (unweighted)}} \right\}$$

For colour television, it is usual to measure the noise in the luminance and chrominance channels separately, through appropriate filters and weighting networks.[2]

Allnatt and Prosser[3] have found that a luminance-weighted s.n.r. of 38·6 dB, and a chrominance-weighted s.n.r. of 32·1 dB, are the minimum values needed to ensure at least 95% favourable* opinions on a colour television picture of above-average sensitivity to noise. However, it must be remembered that the satellite broadcast link is only one of several possible sources of noise. The CCIR recommends[2] that the s.n.r. (luminance-weighted) of a hypothetical reference circuit should not fall below 52 dB for more than 1% of any month. It is reasonable to suggest that the final satellite broadcast link to the viewer should be permitted to have a somewhat lower standard of performance than such a reference circuit, so we will assume that a luminance-weighted s.n.r. of 48 dB is a realistic proposal. The corresponding chrominance-weighted s.n.r. based on the CCIR Recommendation, would be 42 dB. Applying the appropriate weighting factors for flat-spectrum noise and for triangular-spectrum noise (such as would occur in an f.m. system without pre-emphasis), the required s.n.r.'s (unweighted) are given in the following table.

* This means that 95% of the viewers would grade the picture impairment as less than that corresponding to about grade 3·5 on the EBU impairment scale. General experience would indicate that in this case approximately 50% of viewers would assess the impairment as grade 1, and approximately 50% as grade 2 or above.

Unweighted Video Signal-to-noise Ratios Needed to Fulfil Requirements for Colour Television

Noise spectrum	luminance requirement	chrominance requirement
flat	41·5 dB	37·8 dB
triangular	35·7 dB	41·7 dB

Thus, for either flat or triangular spectrum noise, an s.n.r. (unweighted) of about 42 dB is required to fulfil the assumed requirements for colour television.

At the same time, it will be assumed that the unweighted s.n.r. of the audio signal is 60 dB, where the s.n.r. is defined as

$$20 \log_{10} \left\{ \frac{\text{r.m.s. audio voltage at maximum modulation}}{\text{r.m.s. unweighted noise in a 15 kHz bandwidth}} \right\}$$

With the usual C.C.I.T.T. weighting network, this corresponds to a weighted s.n.r. of 54 dB.

3. Frequency modulation

In order to draw conclusions on the benefits to be gained by using digital techniques, it is useful to compare them with some familiar reference system.

Amplitude modulation enables a very simple adaptor to be used at the receiver.[4] However, studies have shown that because of the probable satellite power limitations, and the amount of protection required against co-channel interference, frequency modulation (f.m.) is likely to have significant advantages compared with a.m., without incurring undue complexity at the receiver. Therefore, frequency modulation will be used as the reference against which the digital systems are compared.

It will be assumed that in the f.m. reference system the sound signal is carried by a frequency-modulated 6 MHz sub-carrier added to the composite vision signal, the overall baseband bandwidth occupied then being 6·5 MHz. It has been shown[5] that the subcarrier should produce a peak-to-peak deviation of about 13% of the total deviation, so that 87% is available for the vision signal.

If the effective bandwidth is b_r MHz, it is shown in the appendix that the carrier-to-noise ratio (c.n.r.) at the input, required to produce the given video and sound s.n.r.'s, is given by

$$\text{c.n.r.} = 67{\cdot}3 - 10 \log_{10} [b_r^3 - 26b_r^2 + 169b_r] \quad \text{dB}$$

This relationship is shown in Fig. 1. It is seen that the c.n.r. varies from about 32·7 dB at a bandwidth of 24 MHz to 26·7 dB at a bandwidth of 32 MHz. These figures are well above the f.m. threshold level, so that signal fading is not likely to produce a catastrophic rise in the output noise.

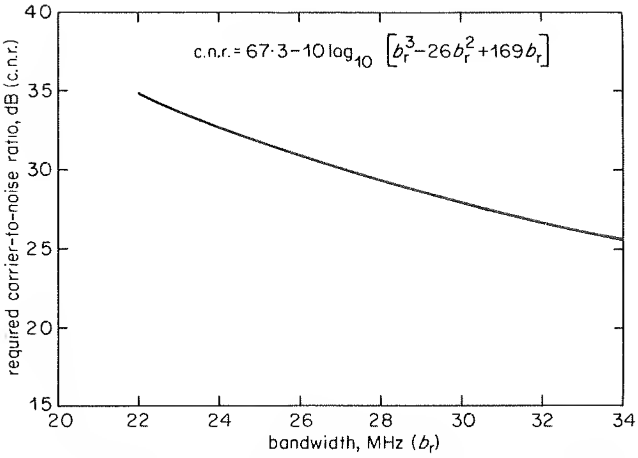


Fig. 1 - Variation of carrier-to-noise ratio with bandwidth, for f.m.

The protection ratio required against co-channel interference is discussed later in Section 6.

Possible effects due to multipath propagation and phase fluctuations are not at present known, but it is believed that they will not be significant in practice, and they will not be considered further in this report.

4. Digitally-coded systems

4.1. Acceptable error rate

When comparing digital with analogue systems it is necessary to establish comparable performance criteria in terms of s.n.r. in the analogue case, and error probability in the digital case. Recent work suggests that the probability of error in interpreting a level of a sampled picture should not exceed about 10^{-7}, if a picture quality equivalent to that assumed in Section 2 is to be obtained. A similar figure will apply to the sound transmission.[6] In considering the sampled error rate, it should be remembered that each sample may be defined by a group of digits, and a distinction must strictly be made between the sample error rate and the digit error rate. However, because the digit error probability varies rapidly with s.n.r., a given sample error rate will require an s.n.r. not very much greater than that corresponding to the same digit error rate.

4.2. Pulse shape and bandwidth

The minimum sampling rate required to define an analogue baseband signal of bandwidth f is 2f, but to allow for practical limitations the actual sampling rate is usually increased by at least 10% to 2·2f. It is assumed that to provide a television picture of sufficiently high quality, the sampled signal must be capable of registering one of at least 128 amplitude levels.[7] In a binary number system, 7 digits are therefore required to define each sample. Future developments may show that a 6-digit system can be made to produce satisfactory results, but this possibility will be discounted for the present. Thus to transmit picture information occupying a bandwidth of 5·5 MHz, the binary-digit rate (bit-rate) must be at least (2·2) (5·5) (7) $\triangleq$ 85 Mb.s^{-1}. The sound signal demands 0·385 Mb.s^{-1}.[8]

In addition, a certain amount of information must be transmitted for purposes of synchronisation, identification, etc. so it may be assumed that provision should be made for a total of about 90 $Mb.s^{-1}$. If this amount of information is transmitted as a binary code, the minimum base-bandwidth occupied cannot be less than 45 MHz. (Practical filters must have a finite response beyond the nominal edge of the pass-band, but for the present purpose it is assumed that the term 'bandwidth' implies a flat response to the edge of the band, and zero response beyond this edge.) Since 45 MHz requires a larger r.f. bandwidth than that available, (see Section 1(c)), a binary system cannot be employed; some form of multilevel pulse must be used.

An important factor in determining the bandwidth demanded by a digital system is the shape of a single pulse. If matched filtering[9] is used at the receiver (i.e., the effective filter transfer characteristic is the complex conjugate of the received pulse spectrum), the s.n.r. obtainable at the receiver output, when random white noise is added in the transmission path, is independent of the received pulse shape. The shape of the transmitted pulse should be chosen so that the energy outside the specified bandwidth is as small as possible, to avoid interfering with adjacent channels. The receiving filter characteristic should then be such that the pulse, after passage through filters at the transmitter and in the receiver, is zero at every sampling interval except that during which the pulse is required to be sampled. Unfortunately, this conflicts with the requirement for a restricted bandwidth, and in practice some compromise between bandwidth and time duration of the pulse must be accepted. This problem has been treated extensively (see, for example, references 10 and 11); in many applications a sine-squared (sometimes called raised-cosine) pulse is a good practical compromise. The bandwidth of the transmission system depends on the amount of interpulse interference which is considered to be acceptable. If no interference between successive pulses is permitted, and if each sample of an analogue signal is represented by a single sine-squared pulse, the base-bandwidth required for a sine-squared pulse is approximately double the analogue signal bandwidth. If the bandwidth is insufficient to permit a sine-squared pulse to be accurately transmitted, either an increased amount of interpulse interference must be accepted, or a pulse shape approaching the theoretically optimum sin x/x type must be used, in which case the instrumentation must meet more stringent performance requirements.

4.3. 4-level single-sideband transmission

If a 4-level (quaternary) system of coding is used, the minimum base-bandwidth will be reduced to 22·5 MHz. This could in principle be transmitted in a single-sideband system within a bandwidth of 24 MHz. If the larger bandwidth of 32 MHz were available, it could be employed to ease the stringent requirements of the pulse shape. However, the problems involved in an s.s.b. system are considerable,[12] and it is unlikely to be practicable for a domestic installation, so it will not be considered further.

4.4. Double-sideband phase-shift keyed system

If double-sideband transmission is to be employed within the restricted bandwidth available, each pulse must be capable of registering one of several values, or 'levels'.

As discussed in Section 4.2, the required information rate is taken as about 90 $Mb.s^{-1}$. Assuming that sin x/x pulses are used, the pulse rate can be as high as the r.f. bandwidth of 24 MHz, thus permitting each sample of the picture signal to be represented by two pulses. There will then be 90/24 = 3·75 information bits per pulse, so that the theoretical number of levels in each pulse would be $2^{3 \cdot 75}$ = 13·5. In practice there must be an integral number of levels, and there are instrumental advantages in having an even number. Furthermore, the number should be as small as possible, to provide better protection against the effect of added noise. By slightly reducing the information rate, and increasing the bandwidth, 12-level pulses can be used. For example, with 12 levels, information can be transmitted at a rate of 89 $Mb.s^{-1}$ in a double-sideband system occupying a bandwidth of 25 MHz. The pulse shape for this bandwidth restriction must approach the sin x/x type, and it would be desirable to ease this stringent requirement by extending the bandwidth to 32 MHz.

It has been shown elsewhere (see, for example, reference 13) that multi-phase modulation has a better noise performance than multi-amplitude modulation, and it can be accommodated within the same bandwidth if the amplitude of the transmitted signal is allowed to vary between the pulse sampling times. Thus, it is not pure phase modulation, but in this report it is referred to as phase-shift-keyed (p.s.k.) modulation.

We shall consider the performance of a 12-phase p.s.k. system for the present application. If the information rate could be reduced, for example, by using 2^6 (instead of 2^7) quantising levels in the picture, an 8-phase system may be feasible.

If an ideal coherent phase detection system is employed, the c.n.r. for a digit error probability of 10^{-7} is approximately 23 dB,[14] and it increases to 26 dB for a differential system, in which each pulse provides a phase reference for the next pulse. A differential system avoids the need for a locked oscillator, and it recovers rapidly from an interruption.

Comparing these results with the f.m. reference discussed in Section 3, we see that if the pulses in the digital system can be maintained at the optimum shape for the 24 MHz bandwidth, the digital system has potentially an advantage of 6·7 to 9·7 dB compared with standard f.m. occupying the same r.f. bandwidth. If, however, the digital bandwidth is increased to 32 MHz, to reduce the stringent pulse shape requirements, this advantage largely disappears. Furthermore, it must be remembered that the figures given assume a phase detector which is accurately aligned to have an infinitely narrow decision threshold. If practical limits are set to this threshold, the required c.n.r. increases.[14] For example, if there is a $\pm 5^{\circ}$ uncertainty in the decision threshold, the c.n.r. required with a coherent detection system is about 3·5 dB greater than the theoretically ideal value.

4.5. Hybrid pulse code modulation

Although the additional instrumentation complexity may be a disadvantage, it is worth considering the benefit

which may be obtained by using a hybrid pulse coding system,[15] in which part of the available bandwidth is employed to transmit an analogue interpolating signal, which prevents the occurrence of quantising noise. Such a system can in principle be very efficient.

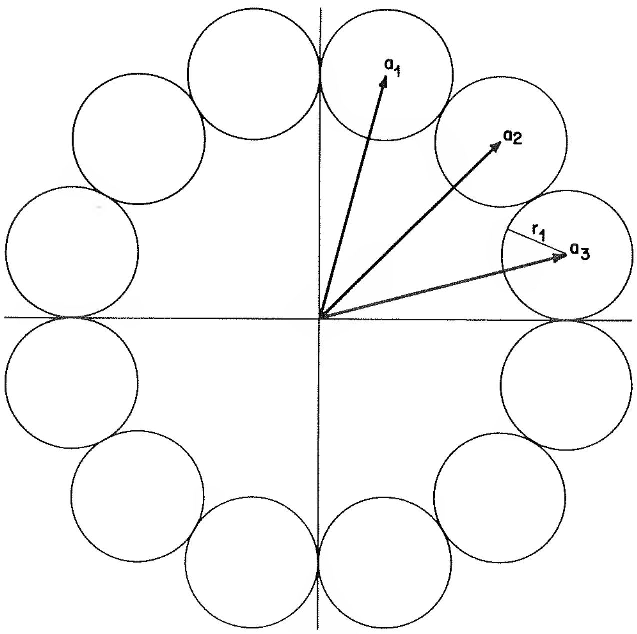
Bearing in mind the requirement of simplicity with respect to filtering and demodulation, a feasible arrangement within the available bandwidth would be one in which one pulse represents the coarsely-quantised signal, and a second pulse is the interpolating signal. Thus, there are two pulses per sample. G.G. Gouriet has suggested that such a two-pulse system could be transmitted efficiently by means of quadrature modulation of a single carrier. In principle, this is an attractive proposal, but it would demand very precise alignment of circuits. A simpler arrangement will therefore be considered for the present requirement. The first pulse could, for example, be transmitted on one carrier in p.s.k. form, and the second by double-sideband amplitude modulation on another carrier. This would appear to be a reasonable compromise between system performance and instrumental simplicity, for the domestic application.

If the system is to have a performance better than that provided by the 12-level p.s.k. system, as discussed in Section 4.4, the peak c.n.r. for a satisfactory picture must be no greater than 23 dB. Since for a pure d.s.b. system the c.n.r. at the peak of modulation for the analogue signal must be about 44 dB, in order to achieve the required demodulated vision s.n.r. (defined in terms of the luminance component) of 42 dB, the improvement in the analogue component given by h.p.c.m. must be at least 21 dB. The number of levels in the quantised component can be no greater than 12, for a satisfactory noise performance with the given 23 dB c.n.r., so the improvement in the analogue component is ($20 \log_{10} 12$) dB, i.e. 21·6 dB.

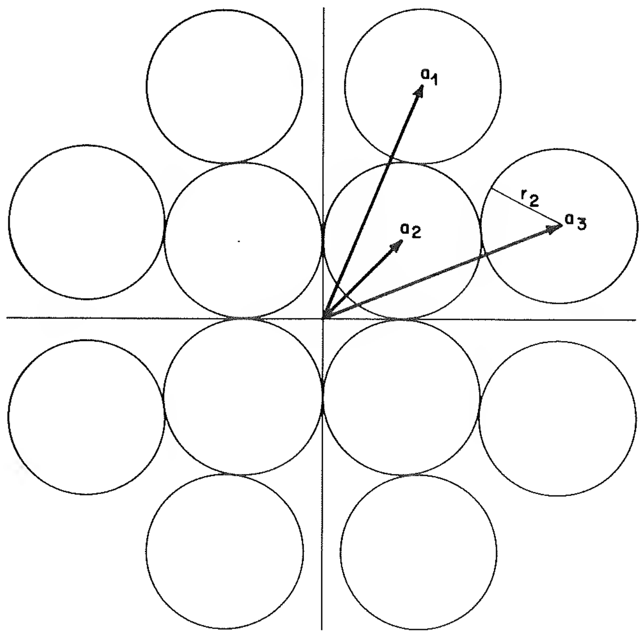
We are thus led to conclude that a practical h.p.c.m. system for a domestic application is unlikely to providea significantly better performance than a normal p.s.k. arrangement, while at the same time the instrumental complexity would be greater.

5. Combined amplitude and phase digital modulation

The 12-phase arrangement discussed in Section 4 represents one of the simpler of the digital detection systems. In order to accommodate such a system within the required bandwidth, the amplitude transfer characteristic of the r.f. transmission system must be linear. This being the case it is reasonable to attempt to make use of this fact by obtaining information from the amplitude as well as the phase of the signal. If this is done, an improved performance in the presence of random noise can be obtained. Various schemes have been proposed,[16,17,18] and formulae have been derived for the error probability in each case, but for the present purpose it is sufficient to take a simple graphical approach to find the order of magnitude of the improvement which can be obtained when it is desired to transmit one of twelve possible signals. Thus, Fig. 2(*a*) shows three possible positions of the carrier vector, a_1, a_2, a_3, in one quadrant of a 12-phase system. The arrow heads indicate the correct positions of the carrier vector; noise is represented by circles such that the resultant of signal plus noise may lie anywhere within the corresponding circle. It is assumed that there will be an error in interpreting the digit value if the circle overlaps with the corresponding circle of an adjacent phase position. Thus the radius r_1 of the circle as drawn provides a measure of the noise voltage which can be tolerated for some specified error probability. In Fig. 2(*b*) the points a_1, a_2, a_3 have been re-arranged, as shown, and the radius r_2 of the circles can now be increased before the circles overlap. This arrangement has 8 phase



(a)



(b)

Fig. 2 - Two arrangements of vectors in a twelve-phase system

positions at the maximum amplitude level and 4 at the lower level. The ratio of r_2 to r_1 is a reasonably good measure of the improvement obtained in the c.n.r. demanded for a given error probability, by using two amplitude levels instead of one. For the case illustrated, the improvement in performance is about 1·7 dB. If the comparison were made on the basis of equal average power, instead of peak power, the improvement would be still greater.

It should be mentioned that this system demands amplitude linearity in the transmitter. Any departure from linearity will increase interpulse interference, and the theoretical noise advantage would therefore be reduced.

6. Co-channel interference

So far in this report comparisons between the systems have been made on the basis of added Gaussian noise. However, the performance of the direct satellite broadcasting system may ultimately be limited more by co-channel interference than by random noise. It is therefore pertinent to compare the systems on this basis.

6.1. F.M.

The protection ratio required between f.m. transmissions of the type discussed in Section 3 has not been completely established, but experiments[5] have indicated that for just perceptible impairment, a ratio of about 30 dB is a reasonable working figure for the bandwidths we are considering. If noise is also present, the effect of adding the subjective impairments of noise and co-channel interference may be regarded as demanding an increase of, say, 3 dB in the protection ratio, which then becomes 33 dB.

6.2. 12-phase keying

If no random noise or distortion is present, a simple approach to this problem is to assume that the interfering signal should be small enough so that when it is added vectorially to the wanted signal, it cannot alter the phase of the resultant vector by $\pi/12$. Thus, the interfering signal must not exceed a relative level of −11·5 dB. This is 18·5 dB less stringent than the corresponding figure for f.m.

In practice, random noise will be present at the same time as the interference. In this case the calculation of the permissible level of interference becomes quite complex. Rosenbaum[19] and Prabhu[20] have considered this problem, and presented some useful curves. Extrapolating some of Prabhu's curves, we find that for a 12-phase coherent detection system with a digit error probability of 10^{-7}, and with a c.n.r. of 26 dB (which is 3 dB greater than the minimum acceptable in the absence of interference, i.e., using the same criterion as in the f.m. case) a single interfering signal level of approximately −23 dB can be tolerated. Less interference can be permitted if there are instrumental deficiencies such as distortion and timing errors. These reduce the immunity of the system to noise and interference. It will be assumed that an allowance of 2 dB must be made to take account of such factors, so that the maximum permissible level of a single interfering signal becomes −25 dB, for a coherent detection system. Rosenbaum concludes that a differential phase-shift detector is likely to be somewhat more sensitive to interference than the coherent system, so that it is reasonable to specify a protection ratio of, say, 28 dB for a practical situation.

The case of combined amplitude and phase digital modulation (as discussed in Section 5) has not been considered in detail, but it is reasonable to assume that the protection ratio required by the combined amplitude and phase system will be no worse than the single-level 12-phase system with coherent detection, and a figure of 25 dB would seem a reasonable proposal.

The preceding discussion has assumed that there is only one interfering signal. If there are a number of simultaneous interferences, the effect upon the wanted signal will tend to be similar to that produced by random noise, and the digital systems will show less advantage over an f.m. system.

7. Interdigit interference

In an f.m. system operating above the threshold level, impairments caused by imperfect circuit responses, multipath propagation and similar defects do not directly affect the noise level, but they may produce other unrelated subjective impairments. In digital systems, however, the error probability depends directly on distortion of the pulse waveform, and the resulting subjective effects may be the same as those produced by noise and interference. It is therefore pertinent to consider the amounts of distortion which may be expected to occur in practice. It is impossible to give exact figures, as these will depend on the initial shape of the pulse waveform. It is useful to assume that the pulse shape will be nearly sine-squared in form, to deduce the probable performance of practical circuits. This is because existing television systems are frequently specified in terms of the K-rating of a sine-squared pulse. The most useful K-rating is based on a pulse whose spectrum is substantially confined within the video band, i.e. the half-amplitude pulse width (2T) is 1/video bandwidth. The interval between such pulses in a digital system is 2T. If we take the maximum of a pulse as the zero time reference, the voltage contributed by this pulse should be zero at time 2T. If the pulse is distorted so that there is some finite voltage at time 2T, this can be considered as interference to the succeeding pulse. It was seen in Section 6 that a co-channel interfering signal of −23 dB may be acceptable in the absence of distortion, with a coherent detection system. It is therefore reasonable to assume that a similar figure may be required for the interpulse interference occurring at the sampling time 2T. This voltage can be given in terms of the K-rating by the relationship

$$\text{voltage at time } 2T = \pm 4K$$

If this voltage is to be not more than −23 dB, K must be not more than 1·8%. This is a fairly stringent specification. Measurements on existing domestic television receivers (for a v.s.b. system) have shown that K may be as high as 6%. However, it should be practicable to achieve a sufficiently good performance in this respect, without undue complexity.

8. Receiver costs

In each of the systems so far considered, it is necessary to demodulate the signal before it is fed (either direct or after re-modulation) to the television receiver. When comparing the systems, therefore, we need only take into account the detecting and decoding circuits. It is extremely difficult to forecast exact costs, but it is nevertheless desirable to estimate tentative figures, even if they are very approximate.

For the f.m. system, two discriminators and some filtering are required, possibly together with a certain amount of amplification. A reasonable estimate of the retail price (on a mass production basis) is taken as £5.

For a digital system, a digital-to-analogue converter, with means of separating the vision and sound signals is required. This is likely to be a relatively complex set of circuits, and the cost will probably not be less than £15. For the phase shift keyed system there is, in addition, the phase detecting arrangement which will include either complex or highly accurate devices, such as a delay line. The costs of the circuitry of various digital detector systems may be roughly comparable, and will probably be not less than £10.

Thus, considering the effect on the retail price, the detection and decoding circuits for a digital system may cost about £25, compared with £5 for the f.m. system. This cost comparison may be considerably affected by future developments in integrated circuit techniques.

9. Conclusions

A practical digital transmission system, operating within the bandwidth which may be allocated to direct television broadcasting from satellites, is unlikely to permit a significantly lower transmitter power than that required for an f.m. system. However, the digital system may have an advantage of about 2 to 8 dB from the point of view of co-channel interference.

If a digital system is employed, a 12-phase p.s.k. arrangement is likely to be the best compromise between performance and complexity. Provided that simple decoding arrangements can be devised, it is worth attempting to accommodate the twelve transmitted phases on two amplitude levels.

It is estimated that the receiver adaptor for a digital system may cost about £20 more than the adaptor for an f.m. system.

Taking the foregoing factors into account there does not seem to be a case for further work to be undertaken at the present time on digital systems for direct broadcasting from satellites. Future developments in digital techniques may, however, make a digital system more attractive than it appears at present, when reliability and economic aspects are included.

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APPENDIX

Derivation of Fig. 1

Let $$D = \left(\frac{\text{peak-to-peak demodulated signal voltage}}{\text{r.m.s. noise voltage after demodulation}}\right)^2$$

$$C = \frac{\text{Carrier power before demodulation}}{\text{noise power before demodulation}}$$

f_d = peak-to-peak deviation, MHz
b_m = bandwidth after detection, MHz
b_r = bandwidth before detection, MHz

Then, $D = 3\left(\frac{f_d^2}{b_m^3}\right) b_r.C.$ (See ref. 14)

i.e. $$C_{dB} = D_{dB} - 10 \log_{10}\left[3\left(\frac{f_d^2}{b_m^3}\right) b_r\right] \qquad (1)$$

If we put $f_d = b_r - 2b_m$, (Carson's rule) and $b_m = 6{\cdot}5$,

Required value of D_{dB} = video s.n.r. + 1·2 dB (to allow for only 87% deviation) + 1·5 dB (to allow for 6·5 MHz bandwidth) + 3 dB (to allow for luminance occupying 70% of peak-to-peak composite signal).

Therefore for a video s.n.r. of 42 dB,

$$D_{dB} = 42 + 1{\cdot}2 + 1{\cdot}5 + 3 = 47{\cdot}7 \text{ dB}$$

$$C_{dB} = 47{\cdot}7 - 10 \log_{10}\left[3\frac{(b_r - 13)^2}{6{\cdot}5^3}\right] \qquad \text{[from (1)]}$$

$$= 47{\cdot}7 + 19{\cdot}6 - 10 \log_{10}[b_r^3 - 26b_r^2 + 169b_r]$$

$$= 67{\cdot}3 - 10 \log_{10}[b_r^3 - 26b_r^2 + 169b_r]$$

DC/SMW/PF

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